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Second Bimonthly Report on the RT-21

Transmitter Development

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I. Purpose

See Bimonthly Report No. 1.

II. Abstract

During the second reporting period, work was divided between a number of areas. As a result of the overall planning of this project, it appears advantageous at the present time to concentrate on the automatic tuning and impedance matching features of the transmitter, while monitoring developments in the high frequency, high power transistor field. No transistor is presently available which will meet the requirements of providing 10 watts at 30 mc. Consequently, in order to evaluate sensing circuitry which ultimately will be used to activate the antenna impedance matching servo system, it has been necessary to design and construct a tunable vacuum tube amplifier capable of simulating the output of the final transistor transmitter. An investigation has been carried out to find a method by which it will be possible to match the transmitter output impedance to the very large antenna impedance range specified, with physically realizable components. Some experimental work has been done to check the operation of the sensing circuitry as well as to measure the voltage dependence of the dielectric constant of a new sample of barium titanate.

The transmitter automatic tuning circuitry reported in the first bimonthly report suffered from a somewhat limited tuning range. During the present reporting period, experiments have been carried out with a tuning circuit which, ideally, has a maximum to minimum frequency ratio equal to the maximum to minimum capacitance ratio rather than the customary square root of the capacitance ratio.

Measurements are included showing the power output capabilities of one

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of the best currently available high power, high frequency transistors. The evaluation of transistors which appear to have characteristics representing an improvement in the high power, high frequency state of the art will continue to be evaluated during the course of this program.

III. Factual Data

1. Transmitter Output Simulator

At the present time, there are no transistors available which are capable of providing an output of 10 watts at frequencies up to 30 mc. It is hoped that by the end of this program, this situation will have improved so that even if a full 10 watts cannot be obtained, the power available will be appreciably closer to the goal. Rather than designing output stages around every high power-high frequency transistor type as it becomes available and discarding them as improved transistors are developed, it has been decided to concentrate effort during the early part of the program on the automatic tuning and matching problems associated with the transmitter.

In order to operate the servo system necessary to accomplish the automatic antenna matching function, it is necessary to design phase and amplitude detector circuits, which will be placed between the transmitter and the antenna. It has been necessary to design and build a vacuum tube amplifier tunable over the 3-30 mc range in order to simulate the transmitter output stage so that the performance of these sensing circuits may be evaluated. Normal frequency generators in the required frequency range do not give sufficient output to operate the sensing circuits correctly. In order to obtain sufficient power to operate the sensing circuits, it would be necessary to make a very unrealistic division of signal power between these

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circuits and the antenna.

Fig. 1 shows a schematic diagram of the tuned amplifier which has been built. This equipment provides an adequate output power level at a wide range of impedance levels. The output impedance is controlled by the selection of the appropriate total capacitance at the output of the pi network. Coarse adjustment is made by means of the selector switch and fine adjustment, by use of the variable capacitor.

Provision is made for operation of the equipment as a crystal-controlled oscillator when accurate frequency checks on the antenna coupler are required. The output may be tuned continuously over the 3-30 mc range by a roller inductor in conjunction with a variable capacitor. The final stage is shunt fed in order to remove D.C. from the output. The driver or oscillator is switched to the appropriate band and peaked for maximum output by the variable capacitor in the plate circuit.

2. Impedance Matching

(i) Introduction

Before the impedance matching problem can be precisely evaluated, the output impedance of the transmitter must be known. Although the 10-watt power transistors are not yet in existence, their output impedance has been assumed to be in the neighborhood of 500 ohms.

The simplest impedance matching device is the L-section. However, a circuit of the form shown in Fig. 2 can only match $R_A > R_O$, while the circuit of Fig. 3 can only match $R_A < R_O$. Since the specified antenna resistance varies between 25 and 1300 ohms, neither circuit is adequate.

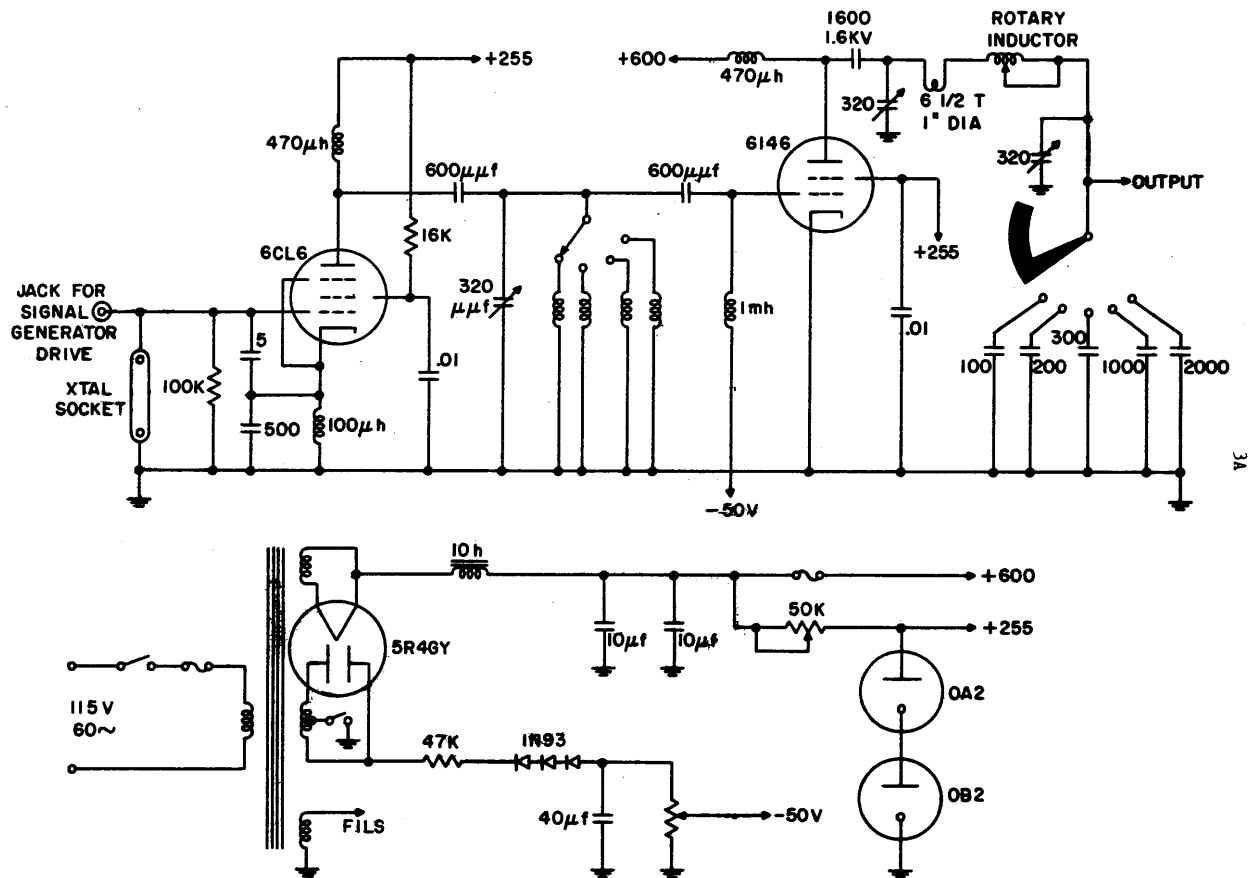


FIGURE 1

3B

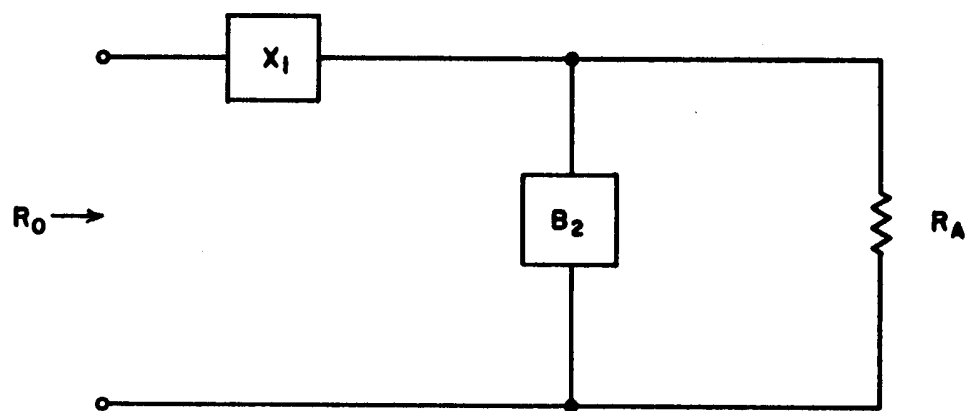


FIGURE 2

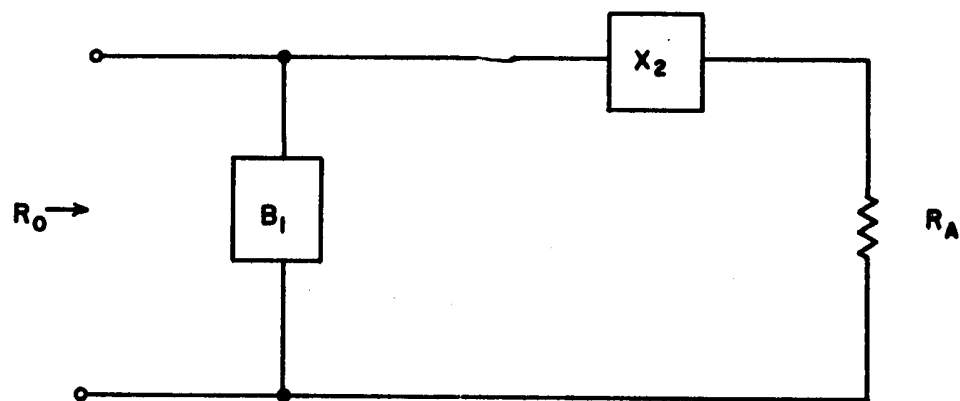


FIGURE 3

- 4 -

The problem of matching any value of antenna impedance to 500 ohms can be theoretically solved with the application of either a tee or pi network. However, the pi network offers the advantage of being able to minimize the effect of any output capacity associated with the output transistor. This minimizing effect occurs if the pi leg shunting the transistor contains a capacitance which is large compared to that of the transistor.

(ii) Pi Matching Network

The design equations of the pi network are simplified if antenna admittance is specified instead of antenna impedance. With the network shown in Fig. 4, the input admittance will be a pure conductance of any specified value G_0 provided

$$B_1 = -B_2 \left[1 \pm \sqrt{\frac{G_0}{G_A} - \frac{G_0^2}{B_2^2}} \right] \quad (1)$$

$$B_3 = -B_2 \left[1 \pm \sqrt{\frac{G_A}{G_0} - \frac{G_A^2}{B_2^2}} \right] - B_A, \quad (2)$$

where G_A is the antenna conductance, and B_A is the antenna susceptance.

The values which may be assumed by G_A and B_A are determined by transforming the impedance plane into the admittance plane. Thus,

$$Y_A = \frac{1}{Z_A}$$

$$G_A + jB_A = \frac{1}{R_A + jX_A}$$

Equating real and imaginary parts leads to

4A

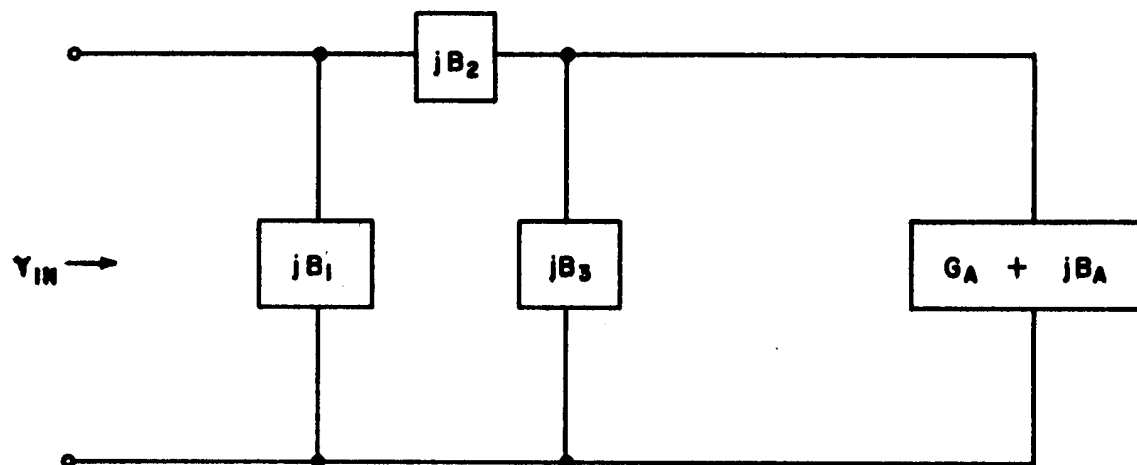


FIGURE 4

- 5 -

$$G_A = \frac{R_A}{R_A^2 + X_A^2}$$

$$B_A = - \frac{X_A}{R_A^2 + X_A^2}$$

Algebraic manipulation of the expressions for G_A and B_A then results in

$$G_A^2 + \left[B_A + \frac{1}{2X_A} \right]^2 = \left[\frac{1}{2X_A} \right]^2 \quad (3)$$

$$\left[G_A - \frac{1}{2R_A} \right]^2 + B_A^2 = \left[\frac{1}{2R_A} \right]^2 \quad (4)$$

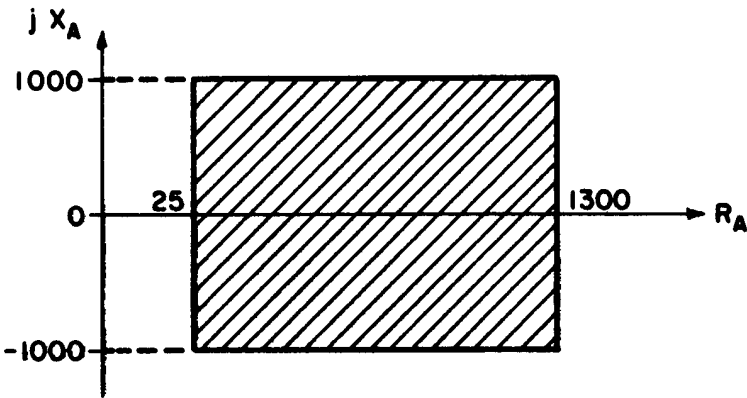
Eq. (3) and Eq. (4) show that lines of constant resistance and lines of constant reactance map into circles in the Y-plane.

The results of mapping the specified antenna impedance into the admittance plane is shown in Fig. 5. The antenna conductance is seen to vary from 25 micromhos to 40,000 micromhos. The susceptance varies between $\pm 20,000$ micromhos. However, an examination of the admittance area indicates that the extreme values of conductance and susceptance do not occur simultaneously.

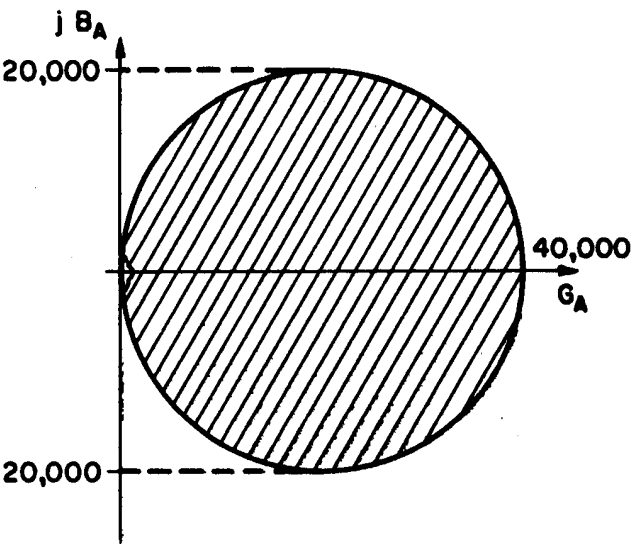
With the variation of antenna admittance known, the variation of susceptance in the shunt legs of the pi may now be evaluated. Since it is desirable for B_1 to be capacitive, B_2 must be inductive. Restating Eq. (1) and Eq. (2) with $-B_2 = 1/X_2$

$$B_1 = \frac{1}{X_2} \left[1 \pm \sqrt{\frac{G_0}{G_A} - G_0^2 X_2^2} \right] \quad (5)$$

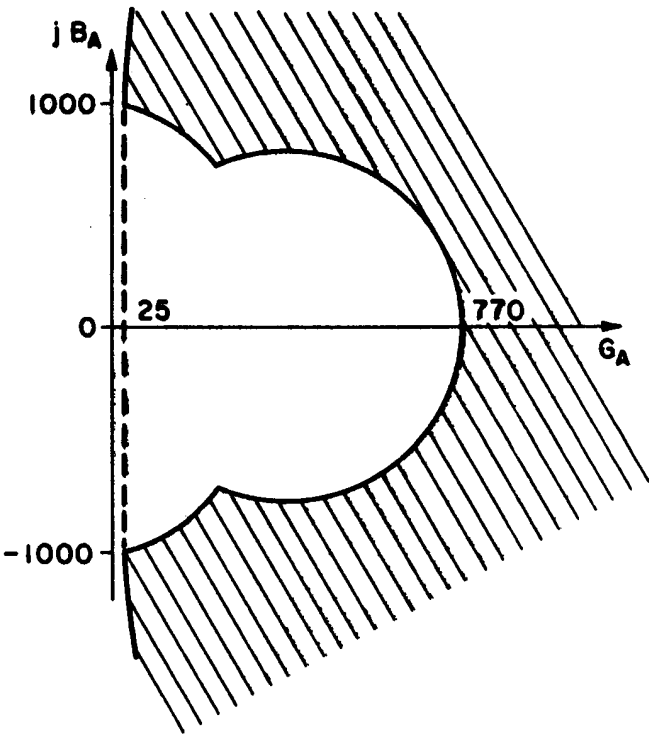
5A



(A) ANTENNA IMPEDANCE (IN OHMS)



(B) ANTENNA ADMITTANCE
(IN MICROMHOS)



(C) ANTENNA ADMITTANCE NEAR ORIGIN
(IN MICROMHOS)

ANTENNA IMPEDANCE - ADMITTANCE RANGE

FIGURE 5

- 6 -

$$B_3 = \frac{1}{X_2} \left[1 \pm \sqrt{\frac{G_A}{G_0} - G_A^2 X_2^2} \right] - B_A \quad (6)$$

Since the quantity within the radical must be non-negative, it follows that

$$\begin{aligned} X_2 &\leq \sqrt{\frac{1}{G_0 G_A}} \\ &\leq \sqrt{\frac{1}{\frac{1}{500} \times 0.04}} = 112 \text{ ohms.} \end{aligned}$$

With X_2 produced by a fixed coil, the restriction that $X_2 = 112$ at 30 mc means that $X_2 = 11.2$ at 3 mc. Analysis of Eq. (5) and Eq. (6) indicates that $B_1)_{\max}$ occurs when

$$\begin{aligned} X_2 &= 11.2 \quad (f = 3 \text{ mc}) \\ G_A &= 25 \text{ } \mu\text{mho} \end{aligned}$$

Then

$$B_1)_{\max} = 0.89 \text{ ohms}$$

$B_1)_{\min}$ occurs when

$$\begin{aligned} X &= 112 \quad (f = 30 \text{ mc}) \\ G_A &= 40,000 \text{ } \mu\text{mho} \end{aligned}$$

Then

$$B_1)_{\min} = 0.0089$$

The range of capacitance required in this leg is then

$$C_1)_{\max} = \frac{B_1)_{\max}}{\omega_{\min}} = 47,400 \text{ } \mu\text{f.}$$

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$$C_1)_{\min} = \frac{B_1)_{\min}}{\omega_{\max}} = 47.4 \mu\text{f.}$$

In a similar manner it was found that

$$C_3)_{\max} = 26,000 \mu\text{f.}$$

$$C_3)_{\min} = 47.4 \mu\text{f.}$$

This pi network which provides a 500 ohm input for all possible antenna admittances over the entire frequency band is shown in Fig. 6.

The circuit of Fig. 6 obviously requires capacitance values which are not very practical. This situation would be alleviated somewhat if two distinct pi networks were employed. One would be used in the 3-15 mc range, while the second would be used when the transmitter was switched to "second harmonic" operation in the 15-30 mc range. With two pi networks, the ratio of maximum to minimum capacitance is reduced by a factor of 4 (to 250:1), but C_1 still requires a rather large maximum capacitance of 23,700 $\mu\text{f.}$

A further improvement would result if, instead of using only a capacitor in the shunt arms of the pi, a circuit were used whose susceptance could be made to vary more rapidly as one of its elements was varied. In the circuit of Fig. 7,

$$B = \frac{\omega C}{1 - \omega^2 LC}$$

$$\frac{\delta B}{\delta C} = \frac{\omega}{1 - \omega^2 LC}$$

while with only a capacitor in the leg

7A

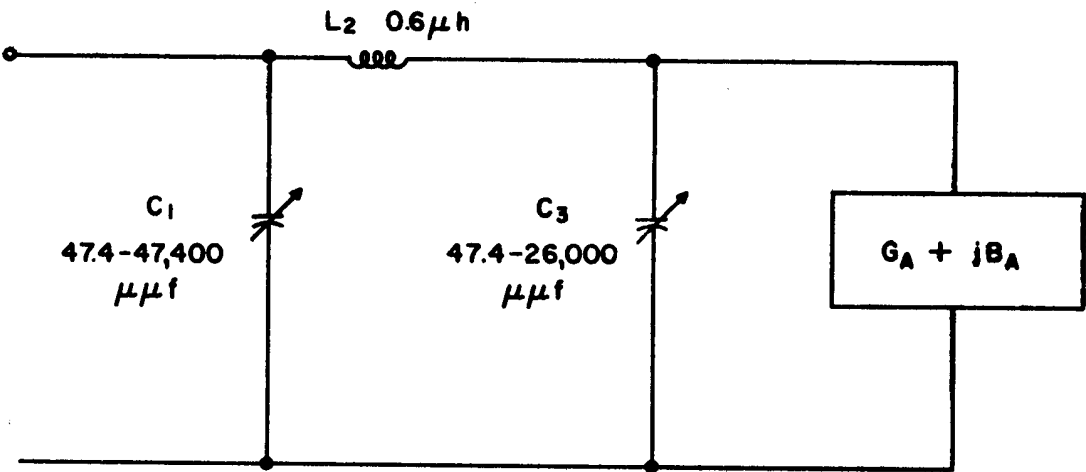


FIGURE 6

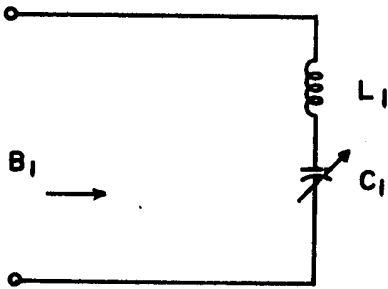


FIGURE 7

- 8 -

$$B = \omega C$$

$$\frac{\delta B}{\delta C} = \omega$$

Thus, with a series LC, not only would the maximum capacitance required be lessened, but maximum to minimum capacitance ratio would also be reduced. Such a circuit is shown in Fig. 8 for the 3-15 mc band. Although this circuit employs elements of a reasonable size, practical difficulties arise. If losses in the coils of the pi legs are taken into account, the admittance of the leg is not a pure susceptance, but rather is

$$Y = \frac{R_L}{R_L^2 + jX^2} - j \frac{X}{R_L^2 + X^2}$$

where

R_L = coil resistance

X = reactance of the leg

When conditions are such that maximum susceptance is required in the input leg of the pi, an equivalent parallel resistance in the neighborhood of 50 ohms is introduced even with a Q of 200 in the coil. This obviously makes a 500 ohm input impedance impossible under these conditions. Actually, due to an interaction among components, a 500 ohm input can be achieved, but losses in the matching network become excessive.

3. Impedance Detector

The circuit shown in Fig. 9 has been built and operated for a load resistance of 500 ohms. Since the desired load impedance has not be definitely established, no extensive test has been given to the impedance detector. However, since it is anticipated that the greatest difficulty will be to induce a phase

8A

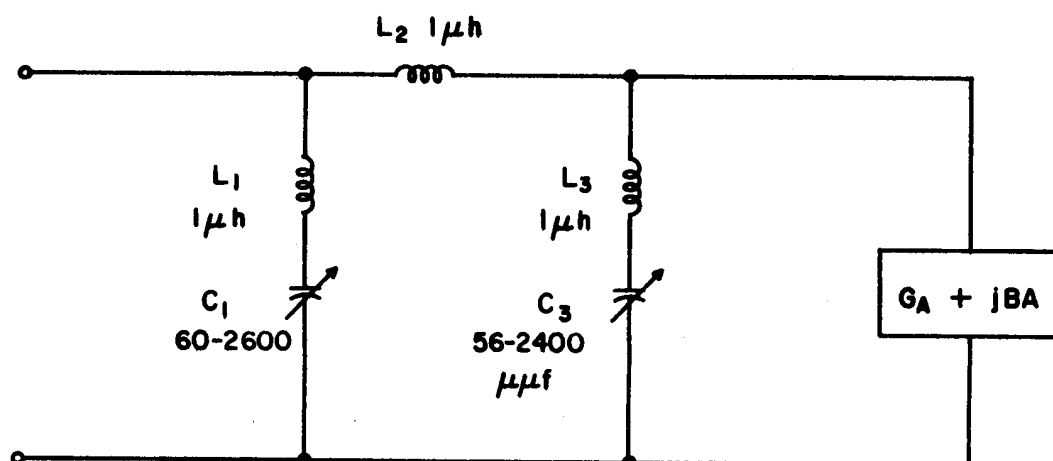
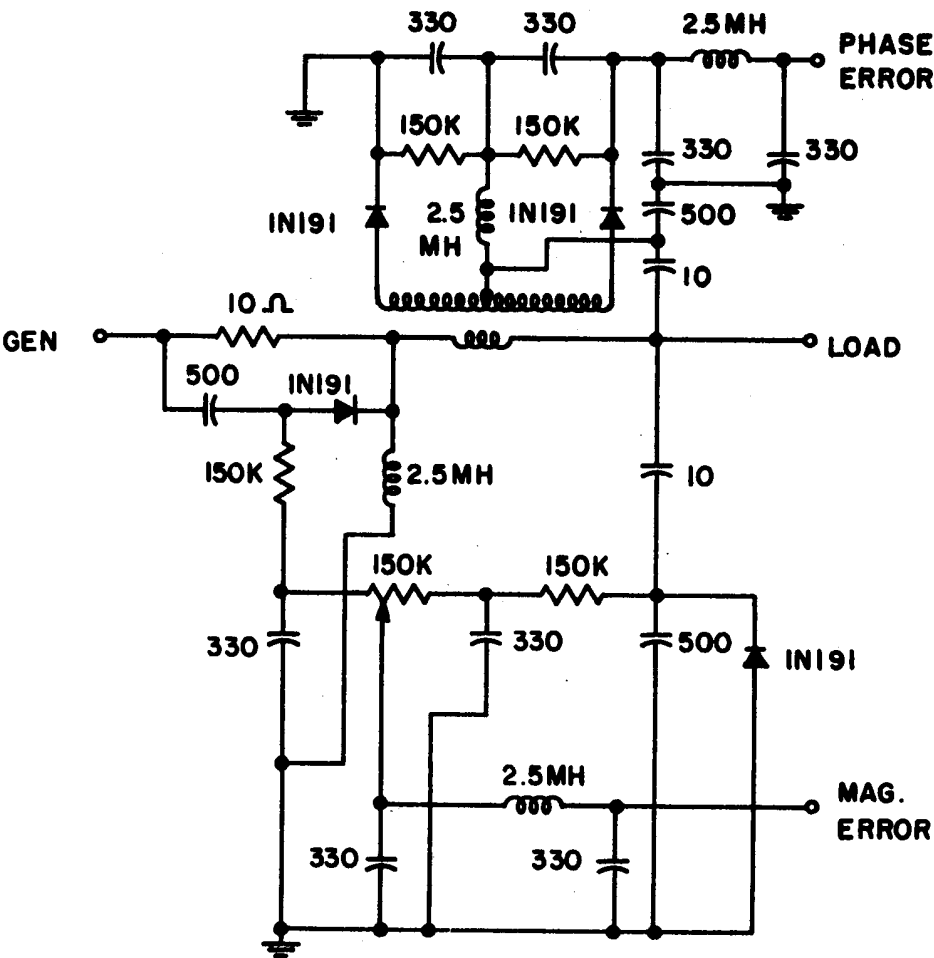


FIGURE 8

8B



IMPEDANCE DETECTOR

FIGURE 9

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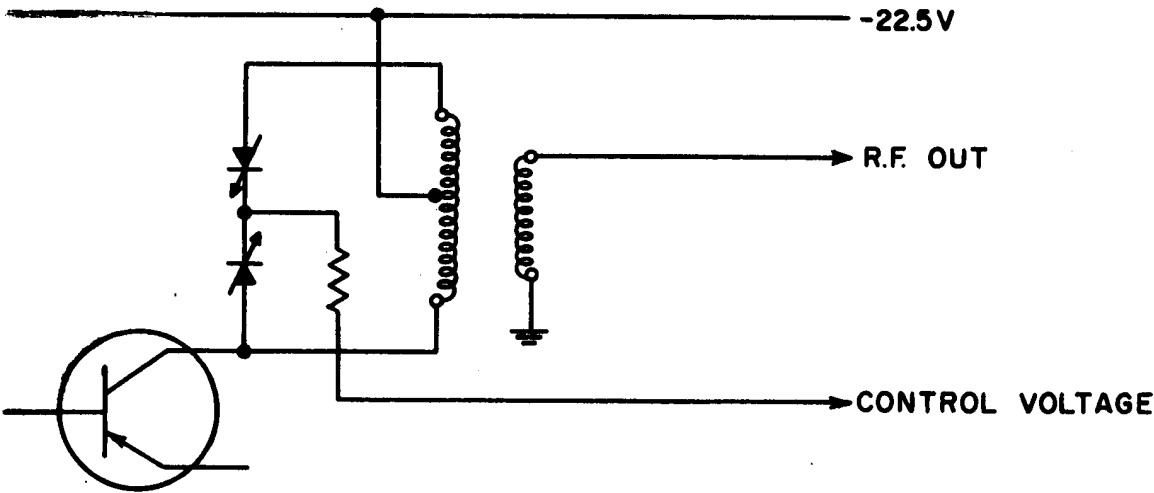
error voltage at the low frequencies, the operation was checked at 3 mc (with 3 watts delivered to the load). The phase error voltage at 3 mc was found to be about 50 mv when the phase angle was 45° . (This phase error, if uncorrected, represents a 1 db. loss.) The polarity of the phase error indicates whether the load is inductive or capacitive, with zero output occurring when the load is resistive. The polarity of the magnitude detector indicates whether the load impedance is larger or smaller than 500 ohms, with zero output occurring when the load equals 500 ohms.

4. Automatic Transmitter Tuning

In the First Bimonthly Report on this program, a system was described by which a limited range of automatic transmitter tuning had been accomplished. Since that time, efforts have been made to increase the frequency range over which continuous tuning could be achieved. Two main approaches have been investigated. Using the original circuit, it became apparent that the range of frequencies being covered was not as great as the maximum to minimum ratio of the voltage variable capacitors would indicate. Investigation showed that although a back-to-back configuration was being used, a self bias was being developed due to the RF currents which were present. This effect became more noticeable in the amplifier stage following the crystal oscillator due to the increased signal level in that stage. In addition to the restriction in tuning range, the self bias resulted in tracking difficulties arising from the differing RF voltages in subsequent stages.

The self bias develops as the result of a lack of symmetry in the tuned circuit. A circuit has been worked out which avoids this difficulty by using center tapped coils as shown in Fig. 10. Work on this circuit was interrupted in order to investigate an arrangement which, under ideal conditions, gives a maximum to minimum frequency

9A



MODIFIED AUTOMATIC TUNING CIRCUIT

FIGURE 10

- 10 -

ratio equal to the maximum to minimum capacitance ratio, rather than the square root of this ratio.

Various configurations of the same basic circuit were studied and for ease of coupling power out of the tuned circuit, the arrangement shown in Fig. 11 was used. In order to evaluate this circuit in as straightforward a manner as possible, a separate, manually tuned oscillator was built. A center tapped coil with an inductance of $3.5 \mu\text{H}$ on either side of the center tap was wound, with physical dimensions as shown in the figure. Using a tuning capacitor with a range of $10 \mu\text{F}$ - $140 \mu\text{F}$ per section, a tuning range from 6 mc to 22 mc was observed without the transistor in the circuit. Resonance was observed with a grid dip meter. With the transistor in the circuit, the maximum frequency fell to 18 mc. Crystals in the range from 6-15 mc were used to test for spurious outputs. It was observed that as long as crystals for frequencies below the lowest limit of the range were not used, no spurious outputs resulted. The low maximum to minimum ratio obtained is attributed to the necessity of coupling power out of the circuit.

Work has been redirected towards increasing the tuning range of the original center tapped circuit by eliminating the self bias mentioned previously.

5. Transistor Evaluation

A short evaluation program was carried out to determine the capabilities of presently available high power, high frequency transistors. The 2N696 transistor, manufactured by Fairchild, was measured as an amplifier at 37.5 mc. This frequency was used purely as a matter of convenience, a test jig being readily available.

It was found that for a substantial power output class A stage, the

10A

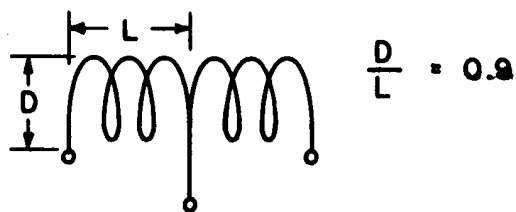
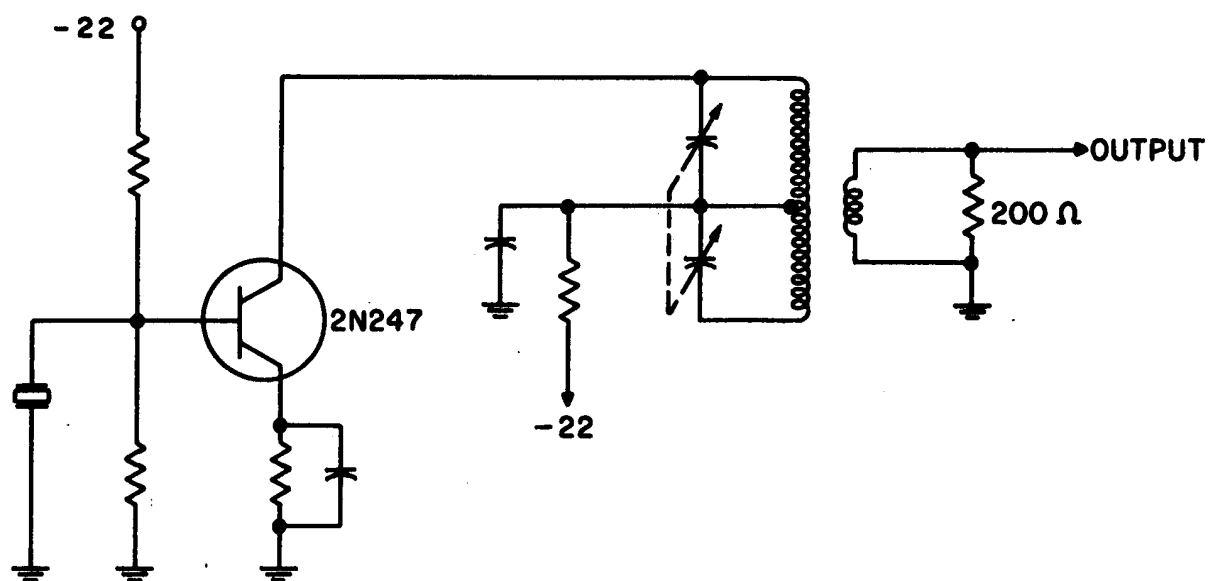


FIGURE II

- 11 -

common base configuration is advantageous, using a load resistance of about 500Ω . A silicon oil heat sink was used. Under these conditions, the case temperature of the transistor exceeds the room temperature appreciably. It was consequently necessary to derate the transistor ($13.5 \text{ mw}/^{\circ}\text{C}$) so that only 1.4 watts were dissipated rather than the 2 watts allowed for a case temperature of 25°C .

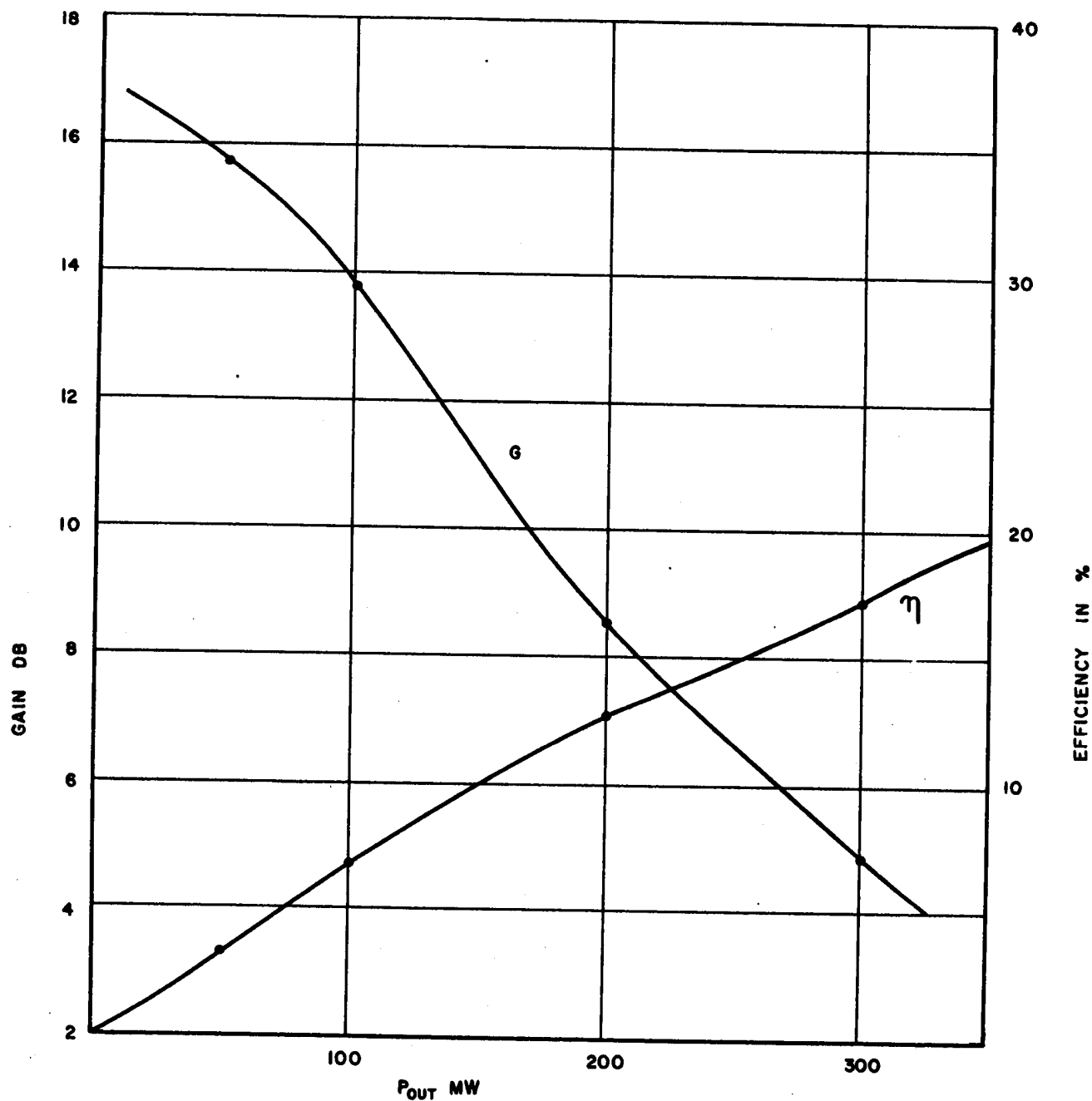
The results are plotted in Fig. 12 which shows the gain and collector efficiency as a function of the output power. The value of V_C was maintained such that $V_{CB}(t)$ did not exceed 40 volts.

Due to the photographic process used in the manufacture of these transistors, units from the same batch are essentially identical. Consequently, these transistors may be connected in load sharing circuits very satisfactorily, the current being shared equally between them.

IV. Conclusions

The construction of the transmitter output simulator has been completed and the equipment tested. No further work is necessary on this unit. It will be used for the evaluation and testing of the impedance detecting circuits. Concerning the antenna impedance matching network problem, the use of a Pi network with variable shunt elements does not appear to present a practical solution. Investigation of other circuit configurations is guided by the postulate that, since the output capacitance of the transistor represents a source of mismatch, it should be made negligible in comparison to a shunting element of the matching network. Also, the matching network should be capable of automatic adjustment by utilizing the detected error in impedance phase and magnitude. At present, the possibility of using cascaded pi-sections is being investigated. The possibility of

11A



GAIN AND COLLECTOR EFFICIENCY OF 2N696
COMM. BASE - CLASS A - STAGE AT 37.5 MC
AS A FUNTION OF OUTPUT POWER

FIGURE 12

- 12 -

incorporating a tuned transformer is also being evaluated. The impedance detecting circuits have been built up and a preliminary check made of their operation. In the absence of a transistor type capable of meeting the output requirements, an estimated figure of 500Ω has been used for the output impedance.

The automatic transmitter tuning work has been continued. An attempt has been made to extend the tuning range by use of a different circuit. In an unloaded condition the tuning ratio of this circuit is equal to the maximum to minimum capacitance ratio. Efforts to couple power out of this circuit have reduced the tuning range until its advantages become marginal. Further investigation of the original circuit have indicated that part of the reason for the restricted tuning range which was observed is due to the development of a self bias developed by the RF signal.

A brief evaluation was carried out to determine the power handling capabilities of the 2N696 transistor. Not surprisingly, this transistor does not come close to meeting the ultimate requirements but does represent the present state of the art. Measurements were also made on some samples of barium titanate for use as voltage variable capacitors. Due either to voids in the material or silver migration of the electrodes, the range of applied voltages was limited by arcing. Consequently, no significant change in dielectric constant was observed.

V. Future Plans

The study of impedance transforming networks will continue. In order to accommodate the large range of antenna impedances required, it may be necessary to transform by means of two cascaded Pi networks. While such a scheme might be designed to handle the full range of impedances, it must be remembered that any

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method must use real components. The more real components, each with their finite losses, which are included in the circuit, the lower the overall efficiency becomes. A breakeven point is reached beyond which the network losses exceed the mismatch losses. An effort will be made to determine this point, assuming components with realizable Q's.

On the basis of information which can now be obtained concerning the output signals from the impedance detecting circuits, a start will be made on the design of the servo amplifiers necessary to control the variable elements in the matching network.

The tuning range of the automatic circuitry is not as great as the maximum to minimum capacitance ratio of the diodes would indicate. Various effects are contributing to this reduction. Each will be studied and attempts made to overcome them. They include self biasing due to the presence of RF voltages and the rapid increase in voltage across the storage capacitor at the start of each sweep. This latter effect results in two side effects, both of which serve to cut down the tuning range. As the voltage on the storage capacitor increases very rapidly, the oscillator does not have sufficient time in which to start oscillation. As the stored charge increases, the frequency sweep is slowed so that oscillation commences; however, due to the slow response of the servo loop, the sweeping action is not stopped before the oscillation ceases. Once the oscillation has ceased, the sweeping action continues to the highest frequency and oscillation cannot start again. At the present time the range of operation is limited to that range of frequencies where the charging rate is slow enough for the servo loop to respond.

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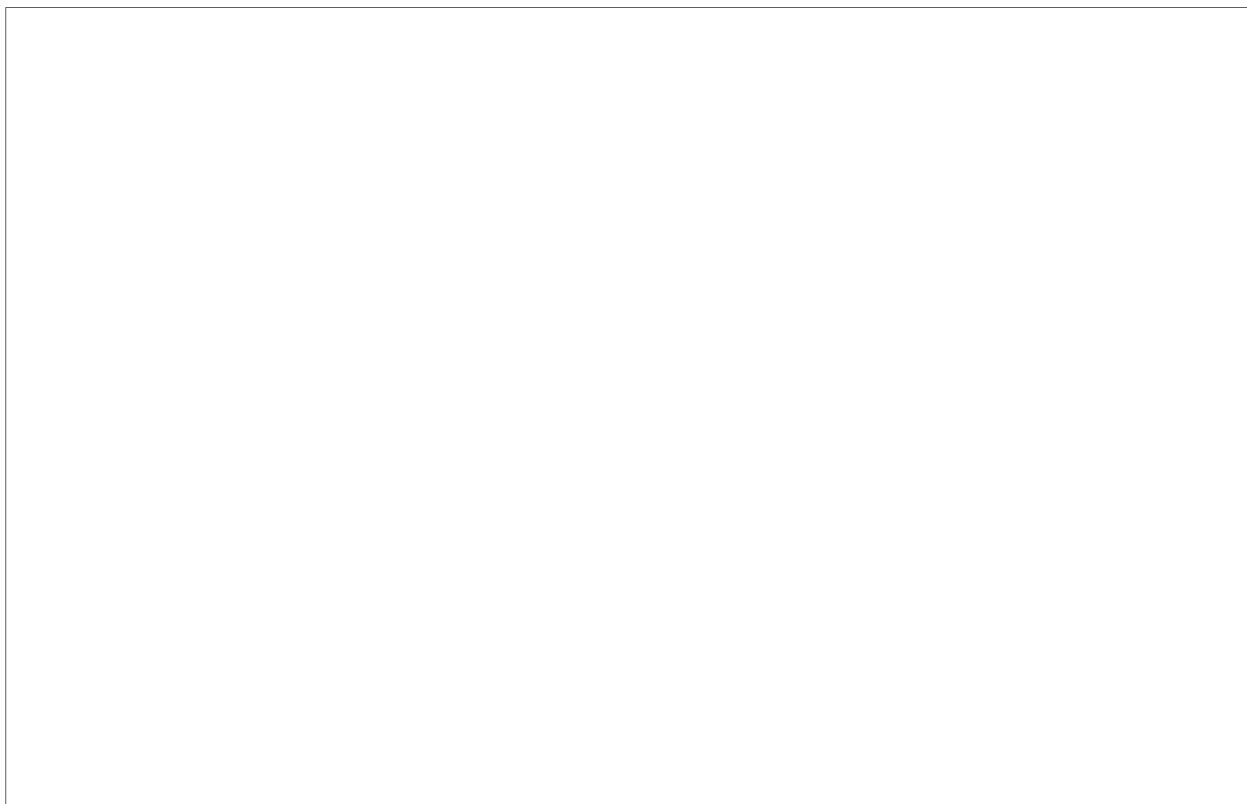
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Transistor evaluation will be continued as and when new types become available.

VI. Identification of Key Technical Personnel

The following name should be added to the personnel reported in the previous bimonthly report.

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